ADAPTIVE FILTERING IN THE PRESENCE OF MULTIPATH

CROSS REFERENCE TO RELATED APPLICATIONS

[0001] The present application relates to and claims priority to applicant's co-pending provisional patent application 60/421,101 entitled: Method and System for Reduced Rank Adaptive Filtering, filed October 25, 2002 which is incorporated herein in its entirety by reference.

[0002] The present application is related to pending utility patent application 09/933,004 entitled System and Method for Adaptive Filtering, filed August 21, 2001, which is incorporated herein in its entirety by reference.

BACKGROUND

[0003] Code Division Multiple Access (CDMA) is a method for multiple users to share common bandwidth over a communications channel. Instead of dividing the frequency spectrum of the channel into discrete sub-channels (frequency division multiplexing – FDM) or alternating use of a channel over time (time division multiplexing- TDM), CDMA mixes each user's transmission with a spreading code, typically a pseudo-random sequence having a starting point and a progression known by both the transmitter and the receiver. In this fashion, many users can share the same bandwidth at the same time. CDMA is one of methods specified for use in third-generation (3G) mobile communications by the International Telecommunication Union,(see, e.g., "The CDMA 2000 Candidate Submission", Telecommunications Industry Association TIA TR-45.5 Subcommittee, June 1998). However, systems employing CDMA are typically interference-limited. The interference can arise from the very nature of the systems in which CDMA is used, e.g., cellular telephone, in which it is typically desirable to accommodate multiple users transmitting simultaneously through a common physical channel.

[0004] In telecommunications, "multipath" is a propagation phenomenon wherein a signal reaches a receiver by two or more paths. Causes of multipath include atmospheric ducting, ionospheric reflection and refraction, and reflection from objects such as mountains and buildings. The effects of multipath include constructive and destructive interference, and phase

shifting of the signal. Interference between signals, e.g., multiple access interference (MAI) can also arise from multiple users sharing a common bandwidth for different services, such as voice and data transmission. Data rate mismatches between users can result in code mismatch. Even in the absence of multipath, this can cause MAI.

[0005] Not surprisingly, since CDMA is currently-specified as a multiple-access scheme for 3G cellular telephone systems and methods mitigating the effects of multipath and providing MAI suppression in CDMA environments are highly desirable. Those systems and methods that are less computation-intensive than others are more desirable.

[0006] Typically, a multiple-delay-line filter, called a "rake," is used to coherently combine multipath signals at the receiver. The phase and gain information used to cohere a rake filter typically is provided by a separate equalization function. Rake receivers, while designed to operate in multipath environments, are generally limited in performance by MAI. MAI causes the performance of a conventional rake receiver to degrade rapidly as the number of users are increased in a CDMA system. On the other hand, research in interference (e.g., MAI) suppression has typically focused on data-aided or blind Minimum Mean Square Error (MMSE) receivers, which are sub-optimal in multipath channels.

SUMMARY

DETAILED DESCRIPTION

[0007] As required, detailed features and embodiments of the invention are disclosed herein. However, it is to be understood that the disclosed features and embodiments are merely exemplary of the invention that may be embodied in various and alternative forms. The figures are not necessarily to scale, and some features may be exaggerated or minimized to show details of particular components. Therefore, specific structural and functional details disclosed herein, and any particular combination of these details, are not to be interpreted as limiting, but merely as a basis for claims and as a representative basis for teaching one skilled in the art to variously employ the invention. This disclosure, while illustrative of preferred embodiments of the invention, describes the invention with respect to Direct Sequence (DS)-CDMA. Preferred embodiments of the invention are also applicable to other modulation types, such as Time

Division Multiple Access (TDMA) and Frequency Division Multiple Access (FDMA), as will be apparent to those skilled in the art. Similarly, although binary signaling is used to illustrated enabling embodiments, embodiments of the invention apply to MAI reduction and multipath mitigation generally.

[0008] In a Direct Sequence (DS)-CDMA system having K users, user k transmits a baseband signal that can be expressed as:

$$x_k(t) = \sum_i A_k b_k(i) s_k(t - iT - \tau_k), \tag{1}$$

[0009] where:

- $b_k(i)$ is the symbol transmitted by user k at i, $b_k(i) \in (-1,+1)$, i.e., binary signaling;
- $s_k(t)$ is the spreading associated with user k;
- A_k is the signal amplitude; and
- τ_k is the delay associated with user k.

[0010] The spreading sequence can be written as:

$$s_k(t) = \sum_{i=1}^{N-1} a_k[i] \Psi(t - iT_c) \tag{2}$$

[0011] where:

- $a_k[i] \in \left(\frac{+1}{\sqrt{N}}, \frac{-1}{\sqrt{N}} \right)$ is a normalization factor for the spreading code;
- $N = \frac{T}{T_c}$ is the processing gain of the CDMA system, or equivalently the bandwidth spreading factor, where T is the symbol period and T_c is the chip period; and
- $\Psi(t)$ represents the chip sequence.

[0012] For the illustrative embodiments described herein, the spreading code is a square wave sequence with no pulse shaping, therefore the chip sequence is a constant. Assume also, without loss of generality, that user 1 is of interest, and that the receiver has timing information to synchronize to the spreading code of the desired user.

[0013] The noiseless distortionless aggregate transmitted signal $\bar{y}(i)$ can be defined as the N-vector composed of asynchronous combination of the data for each user multiplied with its respective spreading sequence. Due to asynchronous transmission, as is inherent to the reverse link of a cellular telephone system, both the current bit of user k, denoted $b_k(i)$, and the previous bit of user k, denoted $b_k(i-1)$ multiplied with their respective portions of the spreading code, are interfering with the current bit of user 1. The noiseless distortionless aggregate transmitted signal can then be written as:

$$\bar{y}(i) = b_1(i)\bar{s}_1 + \sum_{k=2}^{K} A_k\left[b_k(i)\bar{s}_k^+ + b_k(i-1)\bar{s}_k^-\right] \qquad (3).$$

[0014] Here,

- $\bar{s}_1$ is the spreading sequence of User 1;
- $\bar{s}_k^+, \bar{s}_k^-$ are the $N \times (K-1)$ matrices of spreading codes associated with the $K-1$ interfering users.

[0015] Consider the case where the channel is an additive white Gaussian noise (AWGN) channel, so that the distortionless received signal $\bar{r}(i)$ is an N-vector comprising samples at the output of a chip-matched filter at each symbol i, represented by:

$$\bar{r}(i) = \bar{y}(i) + \bar{n}(i) \qquad (4)$$

[0016] Substituting Equation (3) for $\mathbf{y}(i)$, Equation (4) can be written as:

$$\bar{r}(i) = b_1(i)\bar{s}_1 + \sum_{k=2}^{K} A_k\left[b_k(i)\bar{s}_k^+ + b_k(i-1)\bar{s}_k^-\right] + \bar{n}(i) \qquad (5)$$

[0017] It is desirable to extract user 1's baseband signal, $b_1(i)$, while suppressing the interference represented by the summation of Equation (3). Ideally, one would subtract out the interference terms and then despread by user 1's spreading code, $\bar{s}_1$, to extract b_1 from the received signal. In practice, this is difficult because the interference is unknown and embedded in the received signal, and the channel will further distort the transmitted signal, e.g., through the effects of frequency-dependent multipath.

[0018] For an MMSE receiver, receive filter coefficients (denoted by $\bar{c}$) are chosen to minimize the mean-square error (MSE) between the transmitted bit $b_1(i)$ and its estimate. This can be expressed as:

$$MSE = E\left[\left|b_1(i) - \bar{c}^H \bar{r}(i)\right|^2\right] \tag{6}$$

[0019] where:

- $E[\cdot]$ denotes the expected value operator;
- $(\cdot)^H$ represents the Hermitian transpose operator; and
- $\bar{r}(i)$ represents the sample received signal from Equation (5).

[0020] Solving Equation (6) for $\bar{c}$ yields:

$$\bar{c}_{MMSE} = \bar{R}^{-1}\bar{s}_1 \tag{7}$$

where $\bar{R}$ is the covariance matrix given by $\bar{R} = E\left[\bar{r}(i)\bar{r}(i)^H\right]$.

[0021] Note that computation of the MMSE solution for $\bar{c}$ calls for the inversion of the $N \times N$ covariance matrix $\bar{R}$, which can be computationally intense and may not be practical to complete in real time for a high data rate system. Also, if the channel or signals are changing in time, then the sample covariance matrix estimated from the data does not depict the true non-stationary signal environment. Thus, it is desirable to find alternate solutions that approach, or exceed, the performance of the MMSE receiver but that require less processing resources and can adapt rapidly.

[0022] The reduced rank Multistage Wiener Filter (MWF) provides performance comparable to a full-rank MMSE solution, but with much fewer computations and with better performance under lower sample support conditions. An MWF filter achieves a convergence speed-up better than achieved with Principal Components (PC) and reduced computational burden relative to PC. This is done by exploiting information contained in both $\overline{R}_l$ and the steering vector $\overline{s}$ in choosing the reduced dimension subspace that $\overline{s}$ is constrained to lie within. In contrast, PC only makes use of the information contained in $\overline{R}_l$. In MWF, there is no computation of eigenvectors. It has been shown that the MWF implicitly constrains the desired weight vector to lie in the Krylov subspace spanned by $\{\delta_{NM}, \mathbf{K}\delta_{NM}, \ldots, \mathbf{K}^{D-1}\delta_{NM}\}$. MWF is more effective in low sample support environments than other adaptive algorithms. Figure 1 illustrates an MWF with two stages.

[0023] The MWF can be based on a correlation-subtraction architecture as described in "Efficient Architectures for Implementing Adaptive Algorithms" by Ricks and Goldstein, *Proceedings of the 2000 Antenna Applications Symposium*, pg. 29-41. Allerton Park, Monticello, Illinois, September 20-22, 2000 [RICKS], hereby incorporated herein by reference.

[0024] Referring to Figure 1, for an MWF operating in an AWGN channel, $\overline{h}_0 = \overline{s}_1$, i.e., the spreading code of the desired user. Then $\overline{d}_0(n) = \overline{s}_1^H \tilde{x}(n)$, where the input signal $\tilde{x}(n)$ is a vector composed of a block of M bits of the received sampled signal: $[\overline{r}(n-M+1), \overline{r}(n-M+2), \ldots, \overline{r}(n)]$. The filter in Figure 1 demonstrates the low complexity of this MWF implementation and that computation of signal blocking matrices are no longer necessary for subspace partitioning such as that required for constrained adaptation.

[0025] The forward recursion equations, as presented in "A Multistage Representation of the Wiener Filter Based on Orthogonal Projections", Goldstein, Reed, and Scharf, *IEEE Transactions on Information Theory*, Volume 44, Number 7, November 1998 [GOLDSTEIN] and "Adaptive Reduced-Rank Residual Correlation Algorithms for DS-CDMA Interference Suppression" by Honig and Goldstein, *Proceedings of Asilomar*, July 1998 [HONIG 1], hereby incorporated herein by reference in its entirety, along with the initialization and backward recursion equations can be given as:

initialization:

$$d_0(n) \text{ and } \tilde{x}(n) \text{ as above;}$$

forward recursion for $k = 1,2,\ldots,D$:

$$\bar{h}_k = \frac{\sum_\Omega \{d_{k-1}^*(n)\bar{x}_{k-1}(n)\}}{\left\|\sum_\Omega \{d_{k-1}^*(n)\bar{x}_{k-1}(n)\}\right\|},$$

$$d_k(n) = \bar{h}_k^H \bar{x}_{k-1}(n), \text{ and}$$

$$\bar{x}_k(n) = \bar{x}_{k-1}(n) - \bar{h}_k d_k(n); \text{ and}$$

backward recursion for $k = D, D-1,\ldots 1$ with $e_D(n) = d_D(n)$:

$$w_k = \frac{\sum_\Omega \{d_{k-1}^*(n)e_k(n)\}}{\sum_\Omega \{|e_k(n)|^2\}}$$

$$e_{k-1}(n) = d_{k-1}(n) - w_k^* e_k(n)$$

where Ω denotes the region of sample support used to compute the sample statistics.

[0026] In typical applications, the MWF has been limited to environments not overly influenced by multipath, or environments in which multipath correction has been applied at an earlier stage. In these applications, the MWF steering vector is the spreading code of the desired CMDA transmission to be demodulated.

[0027] A channel subjected to multipath interference can be modeled in discrete time by an L-tapped delay line having delay coefficients represented by $\bar{h} = [h_1, h_2, \ldots h_L]$. The parameter L, when viewed in units of time, is known as the delay spread of the channel and is typically on the order of 5-10 microseconds. A signal received via an AWGN channel characterized with multipath in this fashion can be written as:

$$\hat{r}(i) = \hat{y}(i) + \bar{n}(i) \tag{8}$$

where $\hat{y} = \bar{y} * \bar{h}$, and $*$ denotes convolution.

[0028] Substituting for $\bar{y}$ using Equation (3), the received signal can now be written as:

$$\hat{\bar{r}}(i) = b_1(i)\hat{\bar{s}}_1 + \sum_{k=2}^{K} A_k\left[b_k(i)\hat{\bar{s}}_k^+ + b_k(i-1)\hat{\bar{s}}_k^-\right] + \bar{n}(i) \tag{9}$$

where $(\hat{\bullet})$ denotes convolution of the operand with the channel vector $\bar{h}$. Throughout this disclosure, the channel vector is known, or can be estimated in ways known to those skilled in the art.

[0029] Exploiting the analogy between Equation (5) for the AWGN channel and Equation (9) for the multipath channel, the MMSE solution in the presence of multipath can be written from the AWGN solution in Equation (7) as:

$$\hat{\bar{c}}_{MMSE} = \hat{\bar{R}}^{-1}\hat{\bar{s}}_1 = E\left[\hat{\bar{r}}\hat{\bar{r}}^H\right]^{-1}\hat{\bar{s}}_1 \tag{10}$$

[0030] The MWF solution can be obtained similarly, by setting $\bar{h}_0 = \hat{\bar{s}}_1$. Note that the rake-only solution, which incorporates the effects of multipath and neglects the interference, e.g., MAI, can be written as: $\bar{c}_{Rake} = \hat{\bar{s}}_1$.

[0031] In the presence of only AWGN, this reduces to the matched filter (MF) solution, $\bar{c}_{MF} = \bar{s}_1$. The equations above provide a way of representing the Rake, MWF, and MMSE filter coefficients when multipath is present.

[0032] Simulations of the performance of a receiver of the present invention (an multipath MWF receiver) versus a full rank MMSE receiver and a Rake-only receiver indicate that a multipath MWF receiver operating at a reduced rank performs as well as a full rank MMSE receiver and better than a Rake-only receiver.

[0033] The bit estimates can be obtained by multiplying the received signal by the receiver filter coefficients and forming a hard decision. Thus, in the case of MMSE and Rake we compute $\hat{b}_1(i) = \bar{c}_{MMSE/Rake}^H \bar{r}(i)$.

[0034] In the case of a multipath MWF receiver, the bit estimate is produced at the last stage of the filter and can be written as: $\hat{b}_1(i) = \epsilon_0(i)$.

[0035] For the Monte Carlo simulation results, unless otherwise indicated, the processing gain is $N = 32$, and the number of users is fixed at $K = 15$. An $L = 5$ tap channel is used to simulate the multipath, using one tap per chip. The number of bits per block is 5000, and the number of blocks is at least 10 (chosen to produce enough errors to obtain a valid statistical bit error rate estimate). The power of the interfering users is set to 6 dB greater than that of the desired user to determine performance in a near/far situation.

[0036] Figure 2 is a plot of rank versus BER for each receiver type, i.e., full rank MMSE, Rake, and a multipath MWF receiver. The variation in BER for the MMSE and rake receivers is due to the nature of the simulation, since their performance is independent of rank. Note that for rank as low as seven, the simulation indicates that a multipath MWF receiver performs substantially as well as the full rank MMSE receiver and maintains this performance as rank increases. The performance of a multipath MWF receiver degrades only when the rank falls below approximately five. The performance of a multipath MWF receiver at full rank exactly matches the performance of a full-rank MMSE receiver. Both a multipath MWF receiver and an MMSE receiver consistently perform better than the rake receiver.

[0037] Figure 3 is a plot of E_b / N_0 versus BER for each receiver type. In this case, the rank of the multipath MWF receiver is seven. The multipath MWF receiver performs as well as the MMSE receiver and significantly better than the rake receiver. Note that there is about a 6 dB degradation in performance from the ideal Bi-Phase Shift Keying (BPSK) curve. This degradation is directly related to the desired user having 6 dB less power relative to the interferers. To a lesser extent, it is also caused by the system being asynchronous and by the channel characteristics, including multipath. The performance here could be improved by using decision feedback.

[0038] Next, Figure 4 shows a plot of the number of users, K, versus BER for each receiver type. In that Figure, the E_b / N_0 is 15 dB, and the rank of the multipath MWF receiver is again seven. The number of users varies from one to twenty five. These results indicate that for a

lightly loaded system, the Rake receiver performs reasonably well, but its BER degrades rapidly as the load is increased. The multipath MWF receiver substantially meets MMSE receiver performance over the entire range of loading. Note that when the number of users increases beyond half the processing gain, the performance of both MMSE and multipath MWF receivers degrades substantially. This occurs in the presence of asynchronous CDMA signals because the $N+L-1$ taps of both receivers span only one bit interval. A receiver with taps that spans two bit intervals is one way to address this issue. Another way is to increase the sampling rate, which is equivalent to increasing the number of taps per bit. Then, the effect of the asynchronism would have less of an impact.

[0039] In Figure 5, a curve showing the number of bits per block versus BER is shown. For this result, the E_b / N_0 is 15 dB, $K = 12$, and the rank of the multipath MWF receiver is again seven. Note that the MMSE curve requires about 1500 samples to converge to a BER of about $2 \cdot 10^{-3}$. However, the multipath MWF receiver requires only about 1000 samples per block to converge to the same BER. Thus the multipath MWF receiver can be seen as less sensitive to sample support than the full-rank MMSE. This in turn implies that the multipath MWF receiver can track changes in signals that are varying in time faster than MMSE, which is an advantage of reduced rank processing. In addition, the MMSE solution requires inversion of an $N \times N$ covariance matrix, while embodiments of the present invention do not. This illustrates the computational savings that can be obtained by employing the reduced-rank multipath MWF receiver, which yields substantially the same performance as a full rank MMSE receiver. As illustrated in prior figures, the Rake receiver does significantly worse than either the full-rank MMSE receiver or the multipath MWF receiver of the present invention.